(12) EUROPEAN PATENT APPLICATION

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(54) Tracking of sampling frequency in a DAB receiver

(57) A DAB receiver in which, after a DBA baseband signal output from a receiving section is converted into a digital signal at a predetermined sampling frequency, the digital signal is Fourier-transformed by a Fourier transformer to be output has a sampling pulse generator, a DFT window generator serving as a Fourier transformation execution timing, an impulse response arithmetic section for arithmetically operating impulse response of a transmission path on the basis of a Fourier transformation output of a phase reference symbol, an impulse barycenter calculator in the impulse response, and a sampling frequency control means for calculating a change amount of barycentric position on the basis of a position deviation between DFT windows in previous and current frames and barycentric positions in the previous and current frames and controlling the frequency of a sampling pulse on the basis of the change amount of barycentric position.

FIG. 12
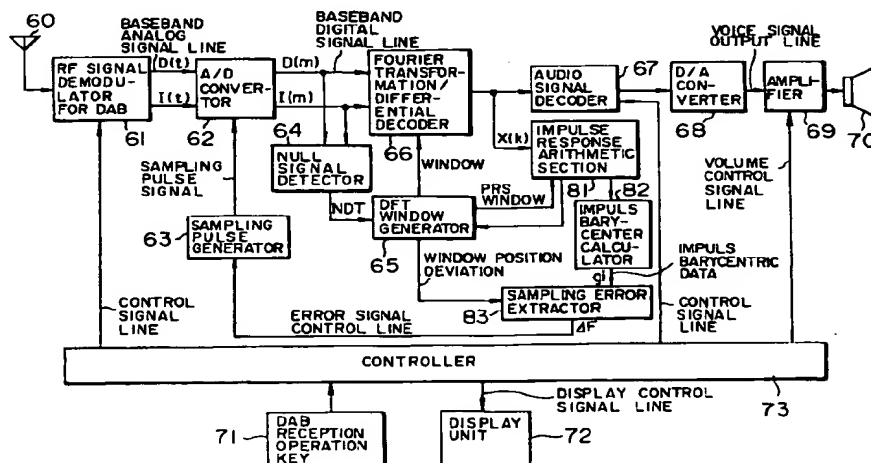
EP 0 836 304 A2

Description

BACKGROUND OF THE INVENTION

5   1. Field of the Invention

The present invention relates to a receiver in digital audio broadcast (Digital Audio Broadcast: DAB) and, more particularly, to a receiver in a digital audio broadcast in which 2N digital data are divided into N groups each having two bits, the first data of each group and the second data of each group are sequentially input to an inverse Fourier transformer as a real-number part and an imaginary-number part, respectively, the real-number part and imaginary-number part output from the inverse Fourier transformer are converted into analog signals, the analog signals are multiplied by cos and sin waves of a carrier frequency fc, respectively, the multiplication results are synthesized to be radiated in a space, signals radiated in the space are received, the received signals are multiplied by the cos and sin waves of the carrier frequency, respectively, the multiplication results are converted into digital signals at a predetermined sampling frequency, the digital signals are input to a Fourier transformer, and a real-number part and an imaginary-number part output from the Fourier transformer are output as the first and second data.

2. Description of the Related Art

20   Digital audio broadcast (DAB) in which audio signals are converted into digital signals to obtain serial data, the serial data are divided into sets each consisting of 2N digital data, the 2N digital data are divided into N groups each having two bits, N carriers having different frequencies depending on combinations of 1 and 0 of each two bits are 4-phase-PSK-modulated, the modulated signals are frequency-multiplexed to be sent from a transmission station, the frequency-multiplexed phase-modulated signal is received by a receiver and demodulated to be an audio output is proposed. The DAB is being studied for practical use in Europe and the like.

In this DAB scheme, in order to reduce influence of selectivity fading, pieces of information are divided in parallel, and modulation is performed by using a plurality of carriers (frequency interleave). As a result, even if any carrier receives fading, influence is reduced as a whole. Basically, the DAB scheme is a frequency division multiplex (FDM: Frequency Division Multiplex) scheme. In a simple FDM, an interval between carriers must be sufficiently large to avoid spectra from overlapping, and good frequency use efficiency cannot be obtained. For this reason, an OFDM (Orthogonal Frequency Division Multiplex) scheme is proposed. In this OFDM, carriers are arranged to satisfy orthogonal conditions and to allow spectra to overlap, good frequency use efficiency can be obtained, and IDFT (Inverse Discrete Fourier Transform) and DFT (Discrete Fourier Transform) operations can be used in a modulator and a demodulator, thereby advantageously obtaining very simple hardware.

(a) Principle of Digital Audio Broadcast by OFDM Scheme

FIG. 13 is a view showing the theoretical arrangement of a transmitter for digital audio broadcast. Reference numeral 1 denotes a serial/parallel converter (S/P converter) for converting serial data d(n) (a(0), b(0), a(1), b(1),...) input at a transmission rate fs (= $2/\Delta t$) into 2N-bit parallel data; 2_0 to 2_{N-1}, N carrier multiplying sections for dividing the 2×N bit parallel data into N groups a(0), b(0),; a(1), b(1);...; a(N-1), b(N-1) each having two bits, multiplying the first bits a(0), a(1),..., a(N-1) of the groups by carriers (cosω_nt) having frequencies f_0 to f_{N-1}, and multiplying the second bits b(0), b(1),..., b(N-1) by carriers (-sinω_nt) having frequencies f_0 to f_{N-1}; and 3, a frequency multiplexer (MUX) for synthesizing output signals a(n)cosω_nt and -b(n)sinω_nt (n = 0 to N-1) from the carrier multiplying sections of the respective groups and frequency-multiplexing the synthesized signal to send a signal D(t).

When output signals from the carrier multiplying sections of the respective groups are synthesized by the frequency multiplexer 3, the carriers having frequencies f_0 to f_{N-1} are 4-phase-PSK-modulated by combinations of 1 and 0 of two bits of each group. The output D(t) from the frequency multiplexer 3 is given by:

50   $$D(t) = \Sigma\{a(n)\cos\omega_n t - b(n)\sin\omega_n t\} \ (n = 0 \ \text{to} \ N-1)$$

Note that, when the frequency interval of carriers is represented by Δf, the (n+1)th carrier frequency fn is given by:

$$fn = f_0 + n\Delta f$$

55

When a transmission time of 2-bit data is represented by Δt (transmission rate fs = $2/\Delta t$), the frequency interval Δf is given by:

$$\Delta f = 1/N\Delta t$$

FIG. 14 is a view for explaining the function of the frequency multiplexer 3. N carriers f_0 to f_{N-1} having intervals Δf are 4ϕ-PSK-modulated by data a(0), b(0); a(1), b(1);...; a(N-1), b(N-1) of N 2-bit groups, and the modulated signals are frequency-multiplexed to be transmitted. FIG. 15 is a view for explaining a symbol. One symbol consists of 2×N bits. When the time length of one symbol is represented by Ts, the following equations can be obtained:

$$Ts = N\Delta t$$

$$\Delta f = 1/Ts$$

Every symbol (2×N bits) is subjected to 4ϕPSK modulation, and the modulated signals are frequency-multiplexed, thereby sequentially transmitting frequency multiplication signals D(t).

FIG. 16 is a view showing a theoretical arrangement of a receiver for digital audio broadcast. Reference numerals 4_0 to 4_{N-1} denote N carrier multiplying sections for multiplying the received signal D(t) by carriers ($\cos\omega_n t$, $-\sin\omega_n t$, n = 0 to N-1) having frequencies f_0 to f_{N-1}; 5_0 to 5_{N-1}, integrators for integrating outputs from the multiplying sections to demodulate data; and 6, a parallel/serial converter (P/S converter) for converting 2×N bit parallel data into serial data.

The integrators 5_0 to 5_{N-1} perform the following arithmetic operation to an input signal D(t):

[Equation 1]

$$\int_0^{\frac{1}{\Delta f} = N\Delta t} D(t)\cos(\omega_m t)dt = \frac{N\Delta t}{2}a(m) = \frac{T_s}{2}a(m)$$

$$\therefore a(m) = \frac{2}{T_s}\int_0^{\frac{1}{\Delta f}} D(t)\cos \omega_m t\, dt$$

to demodulate data a(0), b(0); a(1), b(1);...; a(N-1), b(N-1).

(b) OFDM Modulation Scheme using DFT

In order to generate a baseband signal D(t) of OFDM, N 4ϕPSK modulators are required. In addition, N 4ϕPSK demodulators are required to demodulate the baseband signal D(t). When the number N is large, this scheme is not in practical use. Therefore, a method of simply performing modulation/demodulation by using DFT will be described below.

(b-1) Arrangement of Modulator

A baseband signal of OFDM is expressed by the following equation:

[Equation 2]

$$D(t) = \sum_{k=0}^{N-1} \left\{ a(k)\cos \omega_k t - b(k)\sin \omega_k t \right\}$$

$$(1)$$

When d(k) = a(k) + jb(k) is established, Equation (1) is expressed as follows:

[Equation 3]

$$D(t)=R[D^*(t)]$$

$$=R\left[\sum_{k=0}^{N-1} d(k)\, e^{j\omega_k t}\right] \tag{2}$$

where * means a complex number, and R[] represents the real-number part of [].

Since the following equation is established:

[Equation 4]

$$D^*(t)=\sum_{k=0}^{N-1} d(k)\, e^{j\omega_k t}$$

$$=\sum_{k=0}^{N-1} \left[(a_k+jb_k)\cos\omega_k t + j(a_k+jb_k)\sin\omega_k t\right]$$

$$=\sum_{k=0}^{N-1} (a_k\cos\omega_k t - b_k\sin\omega_k t) + j(a_k\sin\omega_k t + b_k\cos\omega_k t) \tag{3}$$

$$= D(t) + jI(t)$$

it is apparent from Equation (3) that D(t) is expressed by Equation (2). In Equation (2), when t = mΔt is established,

[Equation 5]

$$D(m) = R\left[\sum_{k=0}^{N-1} d(k)\, e^{j\omega_k \cdot m \cdot \Delta t}\right]$$

$$= R\left[\sum_{k=0}^{N-1} d(k)\, e^{j2\pi\frac{km}{N}}\right] \tag{4}$$

where it must receive attention that D(m) is the real-number part of IDFT (Inverse Discrete Fourier Transform) of d(k). When D(m) is output as D(0), D(1),..., D(N-1) for each Δt, D(m) is expressed as shown in FIG. 17. This signal must be obtained by sampling Equation (2) for each Δt. Therefore, when the signal passes through an ideal filter having the frequency characteristic shown in FIG. 18A and the impulse characteristic shown in FIG. 18B, the same equation as Equation (2) can be obtained.

As is apparent from the above description, in the modulator, IDFT is performed to input data d(k) (complex), and the real-number parts D(0) to D(N-1) are output for each Δt and are caused to pass through the ideal filter, thereby obtaining a modulation wave (baseband signal). FIG. 19 is a view showing the arrangement of a main part of a transmitter constituted while giving attention to this point. Reference numeral 11 denotes an IDFT section for performing inverse discrete Fourier transform to input data d(k) which is expressed as complex numbers by data a(k), b(k) (k = 0 to N) of N groups each having two bits; 12a, a DA converter for converting a real-number part output from the IDFT section into an analog signal; 13a, an ideal filter; 14a, a multiplying section for multiplying an ideal filter output D(t) by cosω_ct to perform frequency conversion.

(b-2) Arrangement of Demodulator

When only D(t) is obtained by received frequency conversion, original information cannot be extracted without sampling 2N points. This is because D(t) consists of the real-number part of IDFT at N points. This is apparent from a sampling theorem as shown in FIGS. 20A and 20B. According to the sampling theorem, a signal in a band N · Δf must be sampled at a frequency of 1/(2 · N · Δf). Since the following equation is established:

1/(2 · N · Δf) = Δt/2,

the signal must be sampled at an interval of Δt/2 but an interval of Δt. For this reason, sampling at 2N points must be performed in Ts zone (N · Δt).

However, when a real-number part D(t) and an imaginary-number part I(t) are obtained, an original signal can be extracted by sampling at N points as in the following description.

A complex baseband signal obtained after received frequency conversion is expressed in Equation (5).

[Equation 6]

$$Y^*(t) = \hat{D}(t) + j\hat{I}(t) \tag{5}$$

The right-hand real-number part and the right-hand real-number part are obtained by deforming D(t) and I(t) by the transmission path and noise. When an ideal transmission path is used, the right-hand real-number part and the right-hand real-number part are equal to D(t) and I(t), respectively.

[Equation 7]

$$
\begin{aligned}
Y^*(t) &= \sum_{k=0}^{N-1} (\hat{a}_k \cos 2\pi f_k t - \hat{b}_k \sin 2\pi f_k t) + j(\hat{a}_k \sin 2\pi f_k t + \hat{b}_k \cos 2\pi f_k t) \\
&= \sum_{k=0}^{N-1} (\hat{a}_k + j\hat{b}_k) \cos 2\pi f_k t + j(\hat{a}_k + j\hat{b}_k) \sin 2\pi f_k t \\
&= \sum_{k=0}^{N-1} \hat{d}(k) e^{j\omega_k t}
\end{aligned}
\tag{6}
$$

Assume that Equation (6) is sampled by mΔt (m = 0, 1, 2,... N-1). In this case, when the following equation is established:

[Equation 8]

$$
\begin{aligned}
Y^*(m) &= \sum_{k=0}^{N-1} \hat{d}(k) e^{j\omega_k m\Delta t} \\
&= \sum_{k=0}^{N-1} \hat{d}(k) e^{j2\pi \frac{km}{N}} \\
y^*(m) &= \frac{Y^*(m)}{N}
\end{aligned}
\tag{7}
$$

Equation (7) can be expressed by the following equation:

[Equation 9]

$$y^*(m) = \frac{1}{N} \sum_{k=0}^{N-1} \hat{d}(k) e^{j2\pi \frac{km}{N}} \tag{8}$$

Therefore, d(k) is obtained by Equation (8) as described follows.

[Equation 10]

$$\hat{d}(k) = \sum_{m=0}^{N-1} y^*(m) \, e^{-j2\pi \frac{km}{N}}$$

(9)

According to this equation, an estimation value of the original signal d(k) is obtained. As only a reference, the relational expression between DFT and IDFT is given by the following equation:

[Equation 11]

$$X_k = \sum_{n=0}^{N-1} x_n \, e^{-j2\pi \frac{kn}{N}}$$

$$x_n = \frac{1}{N} \sum_{k=0}^{N-1} X_k \, e^{j2\pi \frac{nk}{N}}$$

(10)

As described above, a complex baseband dignal obtained by frequency-converting a received signal S(t) is converted into a digital signal through a low-pass filter. When the digital signal is subjected to DFT by the DFT section, the estimation value of the original signal d(k) is obtained. FIG. 21 is a view showing the arrangement of a main part of a receiver constituted while giving attention to this point. Reference numeral 15 denotes a frequency converter; 16a and 16b, low-pass filters; 17a and 17b, A/D converters; and 18, a DFT section.

(c) Frequency Conversion on Transmission Side

An orthogonal balance modulation scheme using D(t) and I(t) is the same as a frequency conversion scheme used in an SSB scheme as shown in FIG. 22. Referring to FIG. 22, reference numeral 11 denotes an IDFT section; 12a and 12b, AD converters; 13a and 13, low-pass filters; and 14, a frequency converter. The frequency converter 14 is constituted by multiplying sections 14a and 14b for multiplying $\cos\omega_c t$ and $\sin\omega_c t$ and a hybrid circuit 14c for synthesizing outputs from the multiplying sections to output a synthesized signal.

An output signal S(t) from the frequency converter 14 is given by the following equation:

[Equation 12]

$$s(t) = D(t)\cos\omega_c t + I(t)\sin\omega_c t$$
$$= \sum_{k=0}^{N-1} (a_k\cos 2\pi f_k t - b_k\sin 2\pi f_k t)\cos 2\pi f_c t + \sum_{k=0}^{N-1} (a_k\sin 2\pi f_k t + b_k\cos 2\pi f_k t) \times (-\sin 2\pi f_c t)$$

$$= \sum_{k=0}^{N-1} (a_k\cos 2\pi (f_k + f_c)t - b_k\sin 2\pi (f_k + f_c)t)$$

(11)

The signal S(t) does not include a lower sideband. Therefore, in this scheme, a band which is 1/2 the band in a both-sideband scheme can be obtained, and transmission efficiency is improved.

(d) Frequency Conversion Scheme on Reception Side

FIG. 23 is a view showing the arrangement of an orthogonal frequency converter using $\cos\omega_c t$ and $\sin\omega_c t$. Reference numeral 15a denotes an adder (not actually exist) for adding a noise signal n(t) expressed by the following equation:

$$n(t) = n_c t(k)\cos 2\pi (f_c + f_k)t - n_s(k)\sin 2\pi (f_c + f_k)t$$

6

to a received signal s(t) (see Equation (11)) to output a signal r(t); 15b, a bandpass filter; and 15c and 15d, multiplying sections for multiplying a bandpass filter output by $\cos\omega_c t$ and $-\sin\omega_c t$. In an ideal transmission path which is free from amplitude attenuation, noise, and phase delay, r(t) = s(t) is established. The following description is on the assumption that r(t) = s(t) is established. Output signals D'(t) and I'(t) from the orthogonal frequency converter are expressed by the following equations, respectively:

[Equation 13]

$$D'(t) = \sum_{k=0}^{N-1} [a_k \cos 2\pi (f_c + f_k)t - b_k \sin 2\pi (f_c + f_k)t] \cos 2\pi f_c t$$

$$= \frac{1}{2} \sum_{k=0}^{N-1} [a_k \cos 2\pi f_k t - b_k \sin 2\pi f_k t]$$

$$= \frac{1}{2} D(t) \tag{12}$$

$$I'(t) = \sum_{k=0}^{N-1} [a_k \cos 2\pi (f_c + f_k)t - b_k \sin 2\pi (f_c + f_k)t] \times (-\sin 2\pi f_c t)$$

$$= \frac{1}{2} \sum_{k=0}^{N-1} [a_k \sin 2\pi f_k t + b_k \cos 2\pi f_k t]$$

$$= \frac{1}{2} I(t) \tag{13}$$

Note that, in Equations (12) and (13), the term of $2f_c$ is neglected. Equations (12) and (13) coincide with the real-number and imaginary-number parts of Equation (3) representing a complex baseband signal on the transmission side, respectively. Therefore, as has been described above, an original signal can be extracted by performing an N-point sampling DFT arithmetic operation.

(e) Differential Encoding

Although the frequency conversion scheme on reception side has been described in (d), it is very difficult to form a reception local frequency synchronized with a transmission local frequency ω_c in the OFDM scheme. When the reception local frequency includes a frequency error (asynchronous state), a demodulated vector is rotated as a result, and demodulation cannot be easily performed by the absolute phase. For this reason, information is not represented by the absolute phase, and the information is represented by the magnitude of phase rotation. This is called differential encoding. According to this scheme, even if there is some frequency error, demodulation can be performed.

(e-1) Differential Encoder

FIG. 24 is a view for explaining a differential encoder on transmission side. The logical equation of a differential encoder 21 is given by:

[Equation 14]

$$d_l(k) = \frac{D_l(k) \cdot d_{l-1}(k)}{1 + j} \tag{14}$$

The differential encoder 21 is to convert data $D_L(k)$ which is complex-expressed by the following equations:

[Equation 15]

$$D_l(k) = A_l(k) + jB_l(k) \qquad d_l(k) = a_l(k) + jb_l(k)$$

into $d_L(k)$ by the above logical equation.

A differential code,
in case of (1) $[A_L(k), B_L(k)] = (1,1)$, is given by the following equation:

[Equation 16]

$$D_t(k) = A_t(k) + jB_t(k)$$
$$= 1 + j$$

When this value is put into Equation (14), the following equation can be obtained:

[Equation 17]

$$d_l(k) = \frac{(1+j) \cdot d_{l-1}(k)}{1+j}$$
$$= d_{l-1}(k)$$

the phase does not change.
In case of (2) $[A_L(k), B_L(k)] = (-1,-1)$, the following equation can be obtained:

[Equation 18]

$$D_l(k) = -1 - j$$

$$d_l(k) = \frac{-(1+j) \cdot d_{l-1}(k)}{1+j}$$
$$= -d_{l-1}(k)$$

the phase is inverted, i.e., the phase shifts by π.
In case of (3) $[A_L(k), B_L(k)] = (1,-1)$, the following equation is established:

[Equation 19]

$$D_l(k) = 1 - j$$

$$d_l(k) = \frac{(1-j) \cdot d_{l-1}(k)}{1+j}$$
$$= -j \cdot d_{l-1}(k)$$
$$= e^{-j\frac{\pi}{2}} \cdot d_{l-1}(k)$$

the phase shifts clockwise by $\pi/2$.
In case of (4) $[A_L(k), B_L(k)] = (-1, 1)$, the following equation is established:

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[Equation 20]

$$D_l(k) = 1 - j$$

$$d_l(k) = \frac{(-1 + j) \cdot d_{l-1}(k)}{1 + j}$$

$$= j \cdot d_{l-1}(k)$$

$$= e^{j\frac{\pi}{2}} \cdot d_{l-1}(k)$$

the phase shifts counterclockwise by $\pi/2$.

(e-2) Differential Decoder

On the basis of Equation (14), the following equation is obtained:

[Equation 21]

$$D_l(k) = (1 + j)\frac{d_l(k)}{d_{l-1}(k)}$$

(15)

A differential decoder 22 is to convert data $d_L(k)$ into $D_L(k)$ according to the logical equation represented by Equation (15) as shown in FIG. 25. Therefore, in accordance with the cases (1) to (4) of the differential decoder, the following differential decoding results (1) to (4) are output.

[Equation 22]

(1)

In case of $\dfrac{d_l(k)}{d_{l-1}(k)} = 1$

$$D_l(k) = 1 + j \qquad [A_l(k), \ B_l(k)] = (1, \ 1)$$

(2)

In case of $\dfrac{d_i(k)}{d_{i-1}(k)} = -1$

$$D_i(k) = -1 - j \qquad\qquad [A_i(k),\ B_i(k)] = (-1,\ -1)$$

(3)

In case of $\dfrac{d_i(k)}{d_{i-1}(k)} = e^{j\frac{\pi}{2}}$

$$D_i(k) = j(1 + j)$$
$$= -1 + j \qquad\qquad [A_i(k),\ B_i(k)] = (-1,\ 1)$$

(4)

In case of $\dfrac{d_i(k)}{d_{i-1}(k)} = e^{-j\frac{\pi}{2}}$

$$D_i(k) = -j(1 + j)$$
$$= 1 - j \qquad\qquad [A_i(k),\ B_i(k)] = (1,\ -1)$$

(f) Blocks of Transmission System and Reception System

According to the above description, a transmission system and a reception system in digital audio broadcast using the OFDM scheme have arrangements shown in FIGS. 26A and 26B, respectively. In a frequency converter 14 in the transmission system, reference numeral 14d denotes an oscillator for outputting a cos signal ($\cos\omega_c t$) of a frequency f_c, and reference numeral 14e denotes a phase shifter for shifting the phase of the cos signal by -90° to output $-\sin\omega_c t$. In a frequency converter 15 in the reception system, reference numeral 15e denotes an oscillator for outputting a cos signal ($\cos\omega_c t$) of the frequency f_c, and reference numeral 15f is a phase shifter for shifting the phase of the cos signal by -90° to output $-\sin\omega_c t$. A cos wave and a sin wave (carrier) in the transmission system are called transmission local signals, and a cos wave and a sin wave in the reception system are called reception local signals.

On the transmission side, a DAB frame is constituted by a known phase reference symbol and M data symbols. Each symbol is divided into N groups each having two bits, the first and second data of each group are encoded as a real-number part and an imaginary-number part, respectively. The real-number part and imaginary-number part of a differential code are sequentially input to a Fourier transformer 11, the real-number part and the imaginary-number part output from the Fourier transformer are converted into analog signals, and the analog signals are multiplied by the cos and sin waves of the transmission local frequency fc, respectively. The multiplication results are synthesized to be radiated in the space.

On the transmission side, the signal radiated in the space is received, and the received signal is multiplied by cos and sin waves of a reception local frequency. The multiplication results are converted into digital signals, and the digital signals are input to a Fourier transformer 18. A real-number part and an imaginary-number part output from the Fourier transformer are differentially decoded, and the decoded signals are sequentially output as first and second data serving as original data. The Fourier transformer 18 executes a Fourier transformation process on the basis of the generation timing of a DFT window signal. More specifically, a window signal generator (not shown) detects a null signal portion formed between frames to output a DFT window signal serving as a Fourier transformation execution timing of each symbol. The Fourier transformer 18 executes Fourier transformation on the basis of the generation timing of the DFT window signal.

FIG. 27 is a timing chart for explaining a DAB frame and window signals. The start portion of the DAB frame is called a synchronous channel, and is constituted by a null signal portion NULL and a phase reference symbol PRS

(Phase Reference Symbol). A guard interval GIT of 62 µs is formed after the null signal portion NULL to reduce the influence of a multipath. A predetermined number (= m) of symbols are arranged after the synchronous channel, and a guard interval of 62 µs is formed before each symbol. The contents of the second-half portion of a corresponding symbol are repeatedly inserted into the guard interval.

5   The null signal portion NULL is formed to find the start of a frame. The phase reference symbol PRS is a reference signal for differential decoding and receives a unique pattern (known) inherent in each frame. The PRS window is a window arranged at a predetermined time position with reference to the detected null signal portion NULL. The DFT window indicates the timing of a DFT arithmetic operation of each symbol.

       In the above receiver for digital audio broadcast using the OFDM scheme, the sampling frequencies of AD convert-
10   ers 17a and 17b (FIG. 26) must be made accurately equal to the transmission rate (sampling signal frequency) fs of digital audio data of the transmission side. When the frequencies shift, or a phase difference occurs, the original audio data cannot be correctly demodulated. For this reason, the error amount of the sampling frequency must be monitored, and feedback control must be performed such that the error amount becomes 0.

       However, in a DAB receiver on vehicle, a multipath is generated with movement. For this reason, even if the char-
15   acteristics of the transmission path is calculated by a signal process, the characteristics always change with the passage of time, and the error amount of sampling frequency cannot be easily calculated.

SUMMARY OF THE INVENTION

20   Therefore, it is an object of the present invention to detected an error amount of sampling frequency in the environment wherein a multipath is generated and to perform control to make the error amount zero, thereby demodulating original audio data.

       According to the present invention, the above object is acehived by a receiver in digital audio broadcast using an OFDM scheme, comprising an RF signal demodulator for receiving a signal radiated from a DAB station into the space
25   and multiplying the received signal by cos and sin waves of a carrier frequency to output a baseband signal, an AD converter for converting respective multiplication results into digital data at a predetermined sampling frequency, a DFT window generator for outputting a DFT window signal serving as a Fourier transformation execution timing of each symbol, a Fourier transformer for performing a Fourier transformation process to the digital data for each symbol to demodulate original audio data, an impulse response arithmetic section for arithmetically operating an impulse response of a
30   transmission path on the basis of a Fourier transformation output of the phase reference symbol, an impulse barycentric position arithmetic section for calculating an impulse barycentric position in the DFT window on the basis of the position and level of each impulse included in the impulse response, and sampling frequency control means for calculating a change amount of barycentric position on the basis of a deviation between a DFT window position in a previous frame and a DFT window position in a current frame, an impulse barycentric position in the previous frame, and an impulse
35   barycentric position in the current frame, detecting the frequency error of the sampling pulse on the basis of the change amount of barycentric position, and controlling the frequency of the sampling pulse to make the error zero.

BRIEF DESCRIPTION OF THE DRAWINGS

40   FIG. 1 is a model of a transmission path.
       FIG. 2 is impulse response obtained when no multipath is generated.
       FIG. 3 is a view for explaining a signal in a phase reference symbol.
       FIG. 4 is a graph showing impulse response.
       FIG. 5 is a graph for explaining impulse response depending on the presence/absence of delay.
45   FIG. 6 is impulse response obtained when a sampling frequency has an error in a state wherein no multipath is generated.
       FIG. 7 is an example obtained when impulse response is derived while an FFT window position is controlled in a state wherein a multipath is generated.
       FIG. 8 is a view for explaining FFT window position control in a multipath environment.
50   FIG. 9 is a view for explaining a method of calculating an impulse barycentric position.
       FIG. 10 is the relationship between an FFT window position and a barycenter (when no sampling frequency error exists).
       FIG. 11 is the relationship between the FFT window position and the barycenter (when a sampling frequency error exists).
55   FIG. 12 is the arrangement of a DAB receiver according to the present invention.
       FIG. 13 is a view showing the theoretical arrangement in digital audio broadcast.
       FIG. 14 is a view for explaining the function of a frequency multiplexer.
       FIG. 15 is a view for explaining a symbol.

FIG. 16 is a view showing the theoretical arrangement of a receiver in digital audio broadcast.

FIG. 17 is a view for explaining D(m).

FIG. 18 is the characteristics of an ideal filter.

FIG. 19 is a view showing the arrangement of a main part of a transmitter using IDFT.

FIG. 20 is a view for explaining a sampling theorem.

FIG. 21 is the arrangement of a main part of a receiver using DFT.

FIG. 22 is the arrangement of an orthogonal balance modulation scheme.

FIG. 23 is the arrangement of an orthogonal frequency conversion scheme.

FIG. 24 is a view for explaining a differential encoder.

FIG. 25 is a view for explaining a differential decoder.

FIG. 26 is the arrangements of a transmission system and a reception system.

FIG. 27 is a view for explaining a DAB frame.

DESCRIPTION OF THE PREFERRED EMBODIMENT

(A) Principle of the Present Invention

(a) Shift Amount of Transmission/reception Frame

A DAB receiver specifies a shift amount between a DAB signal frame and a DAB signal frame transmitted by the above method.

FIG. 1 is a model of a transmission path. Reference numeral 51 denotes a DAB signal transmitter; 52, a DAB receiver; and 53, a transmission path. Reference symbol x(n) denotes a transmission signal; h(n), the characteristic of the transmission path, and y(n), a reception signal. The respective signals have the following relationship:

$$X(k) \cdot H(k) = Y(k)$$

i.e.,

$$H(k) = Y(k)/X(k) \tag{16}$$

where X(k), H(k), and Y(k) are obtained by performing Fourier transformation to x(n), h(n), and y(n), respectively.

Due to the above relationship, when x(n) and y(n) or X(k) and Y(k) are known, h(n) (impulse response) is derived, and a shift amount between a signal frame transmitted by the impulse response and a received signal frame can be specified. More specifically, a reference phase symbol (PRS), included in the start of the frame shown in FIG. 27, for capturing synchronization is represented by X(k), and the received PRS is represented by Y(k), thereby deriving h(n).

(b) Relationship between Impulse Response h(n) and Shift Amount of Frames

The relationship between the impulse response h(n) and the shift amount of frames will be described below. When the sampling frequencies of the transmitter 51 and the receiver 52 are equal to each other in the state wherein no multipath is generated, as shown in FIG. 2, an impulse response IPL is observed at a position where n = 0 is established. In FIG. 2, WDW denotes a DFT window. The reason why the impulse response IPL is observed will be described below. In DAB, as X(k), a signal expressed by the following equation is used:

$$X(k) = e^{j\pi^{k/4}} \ (-m \le k \le 1, \ 1 \le k \le m)$$

0 (m does not satisfy the above conditions) This signal is illustrated in FIG. 3 (carrier exists in the range represented by $-m \le k \le -1$ and $1 \le k \le m$). Therefore, the transmission path has no delay time, H(k) is given by the following equation:

$$H(k) = Y(k)/X(k) = 1 \ (-m \le k \le -1, \ 1 \le k \le m)$$

0 (m does not satisfy the above conditions) In the above equation, assume that H(1) = 1 is approximately established. In this case, it is apparent that inverse DFT of H(k) is given by the following sinc function.

$$h(n) = (m/N) \cdot [\sin(n\pi m/N)/(n\pi m/N)] \tag{17}$$

12

More specifically, since m = 1536 and N = 2048 are satisfied in mode I of the DAB, the h(n) is given by h(0) = 1 where n = 0, and the h(n) satisfies h(n) << 1 where n ≠ 0. As shown in FIG. 4, a sharp peak appears at a position where h(n) is maximum, i.e., n (= 0). Therefore, when the transmission path has no delay time, an impulse response shown in FIG. 5A is obtained. However, when the transmission path has a delay time corresponding to n_0 sampling, the impulse response is expressed by h(n - n_0). For this reason, when n is substituted for n - n_0 in Equation (17), the impulse response changes into that shown in FIG. 5B, and an impulse appears at a position which shifts by n_0.

As described above, since the impulse position shifts due to signal delay, the impulse position shifts even if the sampling frequencies of the transmitter and the receiver are not equal to each other. For example, when the sampling frequencies of the transmitter and the receiver are not equal to each other in the state wherein no multipath is generated, an impulse response is observed as shown in FIG. 6.

Therefore, when no multipath is generated, the sampling frequency of the receiver is preferably controlled such that an impulse in the impulse response is generated at a predetermined position of the DFT window, e.g., the position where n = 0 is established.

However, when a multipath is generated, as shown in FIG. 7, a plurality of impulses appear in the impulse response, and the number, positions, and levels of the impulses change with time by a change in multipath environment caused by movement of the receiver. In addition, the DFT window position changes to be slid forward/backward in the symbol zone such that an error rate is minimized depending on the state of the multipath. FIG. 8 is a view for explaining a method of controlling a DFT window position. Reference symbol △ denotes a guard interval, and reference symbol Ts denotes a zone of symbol B. In the guard interval △, the second-half portion of the symbol B is repeatedly inserted. The symbol position of a direct wave received by the receiver is as shown in FIG. 8A, and the symbol positions of reflective wave 1 to reflective wave 3 are as shown in FIGS. 8B to 8D, respectively, because of the delay. As a result, a synthesized wave between the direct wave and each reflective wave is as shown in FIG. 8E. In the synthesized wave, zone a indicates a guard interval zone which is interfered by adjacent symbol A; zone b, a guard interval zone which is not interfered by symbol A; and zone c, the zone of symbol B. A DFT window WDW must be set at a position where each symbol is DFT-demodulated without being influenced by an adjacent symbol. The DFT window WDW is set at a position shown in FIG. 8E.

As described above, in the multipath environment, not only impulses having different positions and different levels exist, but also the position of the DFT window changes. Therefore, the sampling frequency of the A/D converter cannot be easily controlled.

(c) Outline of the Present Invention

(1) When sampling frequency of A/D converter has no error

When a multipath is generated, an impulse indicating a direct wave cannot be easily specified because a plurality of impulses exist. For this reason, the barycentric position of the impulses in the DFT window are calculated, and control is performed on the basis of the barycentric position. As shown in FIG. 9, assume that the start position of the DFT window is represented by 0, that the end position is represented by (N-1), that reference symbol N represents a sample number (number of data of a symbol), that reference symbol i represents a sample position in the DFT window, and that Di represents the amplitude of an impulse at a sample position i. In this case, the barycenter g of the impulses is calculated by the following equation:

[Equation 23]

$$G = \frac{\sum_{i=0}^{N-1} i \cdot D_i}{\sum_{i=0}^{N-1} D_i}$$

(18)

Assume that the sampling frequency of the A/D converter is free from an error in a multipath generation environment. In this case, as shown in FIG. 10A, the barycentric position of the impulses in the DFT window changes by a multipath, and the DFT window position also changes as shown in FIG. 10B. For this reason, assume that the start position of the DFT window (PRS window) of the (L+i)th frame is represented by x_i, that the impulse barycentric position in the DFT window (PRS window) of the (L+i)th frame is represented by g_i, and that the shift amount between the DFT windows (PRS windows) in the (L+i)th frame and the (L+i-1)th frame is represented by p_i. In this case, the change amount

G_i of impulse barycentric position between frames is given by the following equations:

$$p_i = x_{i-1} - x_i \qquad (19)$$

$$G_i = g_i - g_{i-1} - p_i$$

Therefore, a window control amount p_i between frames is given by the following equations:

$$p_1 = x_0 - x_1$$

$$p_2 = x_1 - x_2$$

$$\cdot$$

$$\cdot$$

$$p_L = x_{L-1} - x_L$$

In addition, the barycentric position change amount G_i between frames is given by the following equations:

$$G_1 = g_1 - g_0 - p_1$$

$$G_2 = g_2 - g_1 - p_2$$

$$\cdot$$

$$\cdot$$

$$G_L = g_L - g_{L-1} - p_L$$

An average value D of barycentric position change amounts between L frames is calculated by the following equation:

$$D = \Sigma G_i / L \ (i = 1 \text{ to } L) \qquad (20)$$

By using the average value of the barycentric position change amount, a frequency error amount ΔF is calculated by the following equation:

$$\Delta F = D \times f_{osc} / K \qquad (21)$$

K: number of samples per one frame
f_{osc}: frequency of VCO (sampling frequency).

Therefore, according to (18), (19), (20), and (21), the frequency error ΔF is calculated for each L frames, and the sampling frequency of the A/D converter is controlled such that the frequency error ΔF is 0.

In the case wherein the sampling frequency of the A/D converter has no error as shown in FIG. 10, when the average value D of the barycentric position change values is calculated, the following result is obtained:

$g_0 = 0$

$g_1 = 5,$ $p_1 = x_0 - x_1 = n - (n-5) = 5$
         $G_1 = g_1 - g_0 - p_1 = 5 - 0 - 5 = 0$

$g_2 = -10$ $p_2 = x_1 - x_2 = (n - 5) - (n + 10) = -15$

$$G_2 = g_2 - g_1 - p_1 = -10 - 5 - (-15) = 0$$
$$\therefore D = 0$$

Therefore, it is understood that the frequency error of the VCO for generating the sampling frequency of the A/D converter is 0.

Although the above case is to control a sampling frequency for each L frames, a sampling frequency can also be controlled for each frame. In this case, the frequency error amount ΔF is given by:

$$\Delta F = G_l \times f_{osc}/K \qquad (21)'$$

Therefore, the frequency error amount ΔF of each frame is arithmetically operated on the basis of (18), (19), and (21)' to control the sampling frequency.

(2) When sampling frequency of A/D converter has error

When the sampling frequency of an A/D converter has an error in a multipath generation environment, as shown in FIG. 11A, the barycentric position of impulses in a DFT window changes by a multipath and the frequency error. A DFT window position changes as shown in FIG. 11B.

When the average value D of barycentric position change amounts is calculated as shown in FIG. 11, the following result is obtained:

$$g_0 = 0$$
$$g_1 = 7, \qquad p_1 = x_0 - x_1 = n - (n - 5) = 5$$
$$\qquad\qquad G_1 = g_1 - g_0 - p_1 = 7 - 0 - 5 = 2$$
$$g_2 = -6 \qquad p_2 = x_1 - x_2 = (n - 5) - (n + 10) = -15$$
$$\qquad\qquad G_2 = g_2 - g_1 - p_2 = -6 - 7 - (-15) = 2$$
$$\qquad\qquad \therefore D = 2$$

Therefore, it is understood that a shift amount corresponding to two samples is generated per one frame.

According to the above calculation, a plurality of impulses are generated by a multipath. Even if the position of the DFT window changes, an error component ΔF of the sampling frequency of the A/D converter can be extracted, and the sampling frequency can be controlled such that the error component becomes 0.

(B) Embodiment of the Present Invention

FIG. 12 is a view showing the arrangement of a DAB receiver according to the present invention.

Reference numeral 60 denotes a reception antenna; 61, an RF signal demodulator for DAB which multiplies a received signal by cos and sin waves of a carrier frequency to output baseband analog signals D(t) and I(t); 62, an AD converter for converting the baseband analog signals D(t) and I(t) into digital data D(m) and I(m) at a predetermined sampling frequency; 63, a sampling pulse generator for outputting a sampling pulse; 64, a null signal detector for detecting a null signal portion between DAB frames to output a null detection signal NDL; and 65, a DFT window generator for outputting a DFT window signal and a PRS window signal serving as Fourier transformation execution timings of symbols constituting a DAB frame. The DFT window generator 65 sets windows (DFT window and PRS window) in a zone which is not interfered by an adjacent symbol.

Reference numeral 66 denotes a Fourier transformation/differential decoder for performing a Fourier transformation process to the digital data D(m) and I(m) in the PRS window and the DFT window and differentially decoding the digital data D(m) and I(m) to demodulate X(k) (see Equation (16)) and audio code data (e.g., MPEG audio data); 67, an audio signal decoder for decoding the MPEG audio data into a PCM audio data; 68, a DA converter for converting the PCM audio data into analog data; 69, an amplifier; 70, a loudspeaker; 71, an operation key unit; 72, a display unit; and 73, a controller for performing tuning control or other control.

Reference numeral 81 denotes an impulse response arithmetic section for arithmetically operating an impulse response by using a Fourier transformation output X(k) in a PRS window period on the basis of Equation (16); 82, an impulse barycenter arithmetic section for arithmetically operating an impulse barycentric position gi on the basis of Equation (18); and 83, a sampling error extractor for arithmetic operating an error amount ΔF of the sampling frequency on the basis of Equations (19) to (21) to output the error amount ΔF.

The RF signal demodulator 61 for DAB multiplies a received signal by cos and sin waves of the carrier frequency to output baseband analog signals D(t) and I(t), and the AD converter 62 converts baseband analog signals D(t) and I(t) into digital data D(m) and I(m) on the basis of a sampling pulse output from the sampling pulse generator 63 to input

the digital data D(m) and I(m) to the null signal detector 64 and the Fourier transformation/differential decoder 66. The null signal detector 64 detects a null signal portion between DAB frames to output a null detection signal NDT. When the null detection signal NDT is input, the DFT window generator 65 moves a window to a position which is not interfered by an adjacent symbol depending on a multipath situation to input the window to the Fourier transformation/differential decoder 66, and also inputs the PRS window to the impulse response arithmetic section 81.

When the Fourier transformation/differential decoder 66 performs a Fourier transformation process to the digital data D(m) and I(m) in the PRS window and the DFT window for each symbol, and then differentially decodes the digital data D(m) and I(m) to demodulate X(k) and audio code data (MPEG audio data) and to output them. The audio signal decoder 67 decodes the MPEG audio data into PCM audio data, and the DA converter 68 converts the PCM audio data into analog data and inputs the analog data to the loudspeaker 70 through the amplifier 69, thereby outputting DAB voice.

The impulse response arithmetic section 81 arithmetically operates impulse response by using the Fourier transformation output in the PRS window period on the basis of Equation (16), and inputs the arithmetic result to the DFT window generator 65 and the impulse barycenter calculator 82. The DFT window generator 65 moves a window to a position which is not interfered by an adjacent symbol in the next DAB frame on the basis of the impulse response (multipath).

The impulse barycenter calculator 82 arithmetically operates an impulse barycentric position gi on the basis of Equation (18), and the sampling pulse extractor 83 arithmetically operates the average error amount ΔF of the sampling frequencies in L frame periods on the basis of Equations (19) to (21) to input the average error amount ΔF to the sampling pulse generator 63. The sampling pulse generator 63 controls the sampling frequencies such that the average error amount ΔF becomes 0, and inputs a sampling pulse to the AD converter 62.

Subsequently, the above operation is repeated, the sampling frequency of the transmitter is equal to the sampling frequency of the receiver, and preferable audio voice can be output from the loudspeaker even in a multipath generation environment.

In the above description, the sampling frequencies are controlled on the basis of the average error amount ΔF of the sampling frequencies in the L frame periods such that the error amount becomes 0. However, a sampling frequency can also be controlled for each frame. More specifically, the sampling pulse extractor 83 can also be constituted as follows. That is, the sampling pulse extractor 83 arithmetically operates the sampling frequency error amount ΔF for each frame on the basis of Equations (18), (19), and (21)' and inputs the error amount to the sampling pulse generator 63 to control the sampling frequencies.

Although the present invention has been described above with reference to the embodiment, the various changes of the present invention can be effected according to the spirit and scope of the invention described in the claims. The present invention does not exclude these changes.

According to the present invention, the barycenter of impulse response is calculated for each signal frame of DAB, and an error component of a sampling frequency is extracted in consideration of the position of a DFT window, and the sampling frequency is controlled such that the error becomes 0. For this reason, the sampling frequency of a DAB receiver can be made equal to the sampling frequency of a transmitter even in a multipath environment, and preferable DAB voice can be output.

Claims

1. A receiver in digital audio broadcast in which a frame is constituted by one known phase reference symbol and M data symbols, 2N digital data constituting the data symbols are divided into N groups each having two bits, the first data of each group and the second data of each group are sequentially input to an inverse Fourier transformer as a real-number part and an imaginary-number part, respectively, the real-number part and imaginary-number part output from said inverse Fourier transformer are converted into analog signals, the analog signals are multiplied by a cos wave and a sin wave of a carrier frequency fc, the multiplication results are synthesized to be radiated in a space, signals radiated in the space are received, the received signals are multiplied by the cos wave and the sin wave of the carrier frequency, the multiplication results are converted into digital signals at a predetermined sampling frequency by an AD converter (62), the digital signals are input to a Fourier transformer (66), and a real-number part and an imaginary-number part output from said Fourier transformer (66) are output as the first and second data,

characterized by comprising:

a sampling pulse generator (63) for generating a sampling pulse to input the sampling pulse to said AD converter;
a DFT window generator (65) for outputting a DFT window signal serving as a Fourier transformation execution timing of each symbol;

an impulse response arithmetic section (81) for arithmetically operating an impulse response of a transmission path on the basis of a Fourier transformation output of the phase reference symbol;

an impulse barycentric position arithmetic section (82) for calculating an impulse barycentric position in the DFT window on the basis of the position and level of each impulse included in the impulse response; and

sampling frequency control means (83) for calculating a change amount of barycentric position on the basis of a deviation between a DFT window position in a previous frame and a DFT window position in a current frame, an impulse barycentric position in the previous frame, and an impulse barycentric position in the current frame, and controlling the frequency of the sampling pulse on the basis of the change amount of barycentric position.

2. A digital audio broadcast receiver according to claim 1, characterized in that said sampling pulse generator (63) controls the sampling frequency such that an average error amount becomes 0, thereby inputting a sampling pulse to the AD converter (62).

3. A digital audio broadcast receiver according to claim 1, characterized in that said DFT window generator (65) sets a predetermined window at a position which is not interfered by an adjacent symbol in the next frame on the basis of the impulse response.

4. A digital audio broadcast receiver according to claim 3, characterized in that the predetermined window is a DFT window and a PRS window.

5. A digital audio broadcast receiver according to claim 3, characterized in that the predetermined window is set at the position which is not interfered by the adjacent symbol when a null signal is input.

6. A digital audio broadcast receiver according to claim 5, characterized in that the null detection signal is obtained such that a null signal portion between frames is detected by a null signal detector (64).

7. A digital audio broadcast receiver according to claim 1, characterized in that the impulse response indicates a multipath situation.

8. A digital audio broadcast receiver according to claim 1, characterized in that the phase reference symbol is a PRS window period.

9. A receiver in digital audio broadcast according to claim 1, characterized in that said sampling frequency control means (83) controls a sampling frequency on the basis of an average change amount of barycentric position of L frames.

10. A digital audio broadcast receiver according to claim 9, characterized in that said sampling frequency control means (83) calculates an error amount of sampling frequency by a predetermined arithmetic operation process.
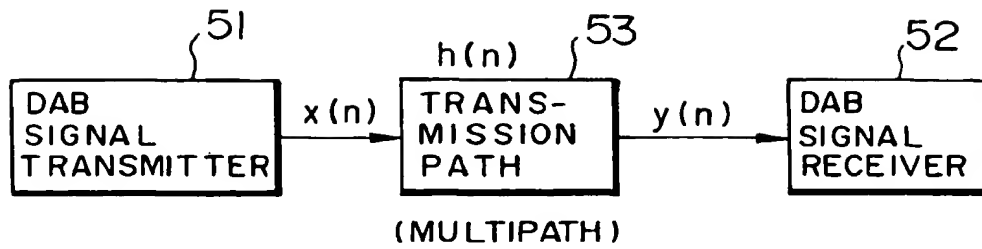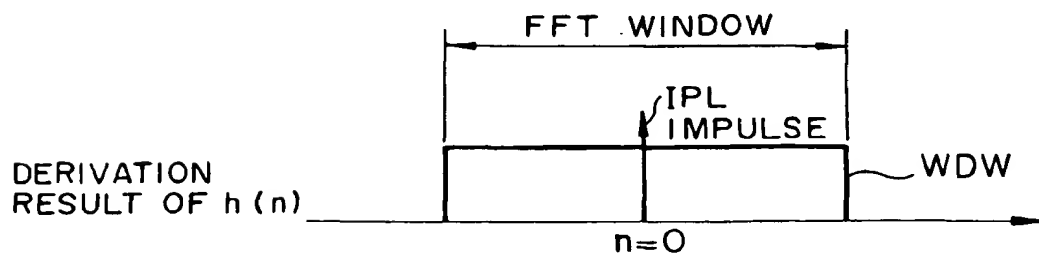
FIG. I

(MULTIPATH)
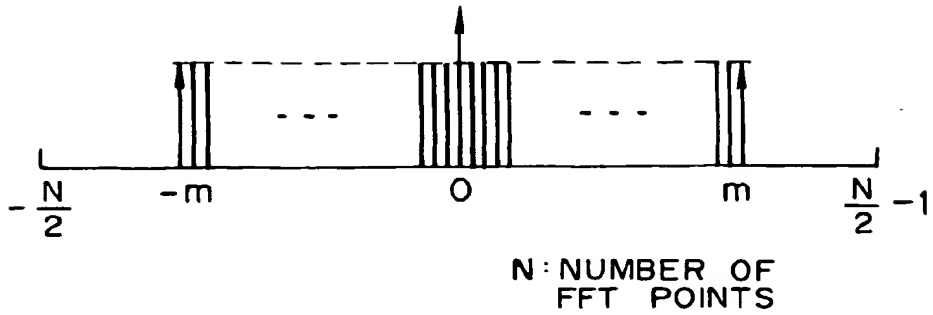
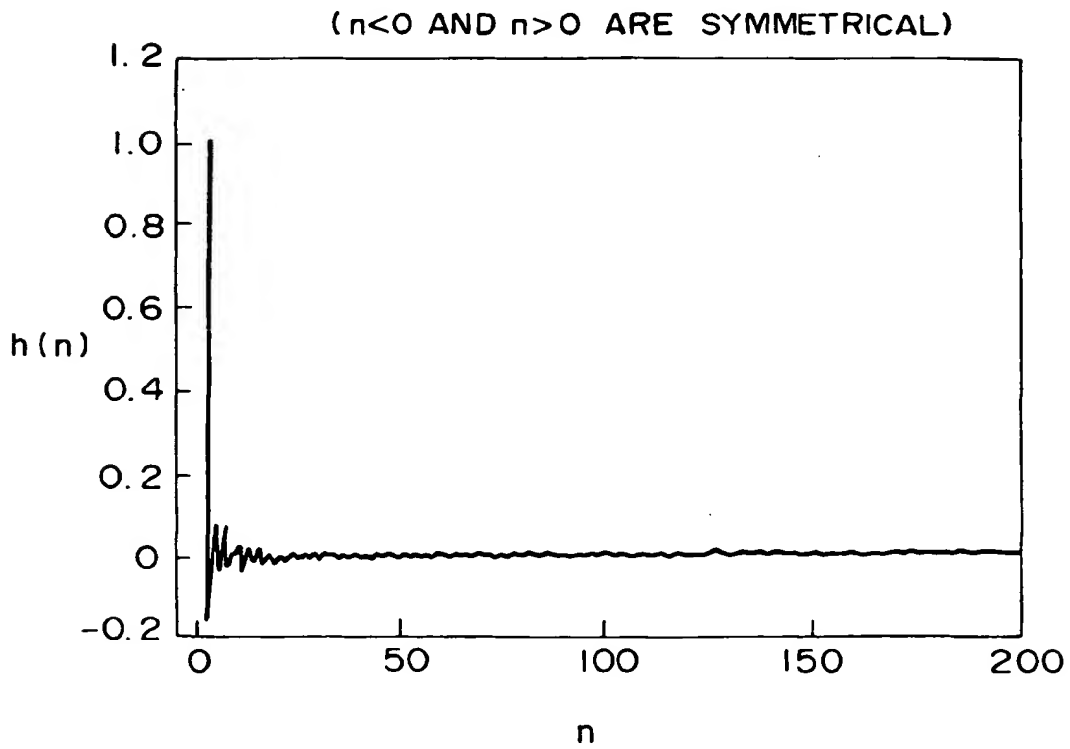# FIG. 2

FIG. 3

$$-\frac{N}{2} \quad -m \quad 0 \quad m \quad \frac{N}{2}-1$$

N : NUMBER OF
FFT POINTS

FIG. 4

(n<0 AND n>0 ARE SYMMETRICAL)

FIG. 5A

FIG. 5B

FIG. 6

FFT WINDOW

DERIVATION RESULT OF
h(n) OF Lth FRAME

O

DERIVATION RESULT OF
h(n) OF (L+1)th FRAME

DERIVATION RESULT OF
h(n) OF (L+2)th FRAME

FIG. 7

FFT WINDOW

DERIVATION RESULT OF
h(n) OF Lth FRAME

DERIVATION RESULT OF
h(n) OF (L+1)th FRAME

DERIVATION RESULT OF
h(n) OF (L+2)th FRAME

FIG. 8

(a) DIRECT WAVE

(b) REFLECTIVE WAVE 1

(c) REFLECTIVE WAVE 2

(d) REFLECTIVE WAVE 3

(e) SYNTHESIZED WAVE

: INTER-SYMBOL INTERFERENCE

FIG.9

FFT WINDOW

i = 0　　　　　　i = N−1

$$G = \frac{\sum\limits_{i=0}^{N-1} i \cdot D_i}{\sum\limits_{i=0}^{N-1} D_i}$$

G : BARYCENTER
i : SAMPLE POSITION OF FFT WINDOW
Di : AMPLITUDE OF IMPULSE
N : NUMBER OF FFT SAMPLES

FIG.IOB

FIG.IOA

FIG.11B

BARYCENTRIC
POSITION

FFT WINDOW

$x_0 = n$ m 2

FFT WINDOW

$x_1 = n-5$ $m+2$ 2

P_1

FFT WINDOW

$x_2 = n-10$ $m+4$

P_2

FIG.11A

FFT WINDOW

BARYCENTRIC
POSITION

$g_0 = 0$

FRAME L

$g_1 = 5+2$

FRAME (L+1)

$g_2 = -10+2+2$

FRAME (L+2)

FIG. 12

FIG. 13

SERIAL DATA : d(n)

S/P CONVER-TER

a(0), b(0), a(1), b(1) ....

$$f_s = \frac{2}{\Delta t} \left(\begin{array}{c} \text{TRANS-} \\ \text{MISSION} \\ \text{RATE} \end{array} \right)$$

a(0) b(0) a(1) b(1) ....

a(0) b(0) a(1) b(1) ....

$\longrightarrow t$

Δt

a(0)
b(0)

a(N-1)
b(N-1)

$\cos \omega_0 t$ EQUAL TO 4ϕPSK

$-\sin \omega_0 t$

$\cos \omega_{N-1} t$

$-\sin \omega_{N-1} t$

MUX

D(L)
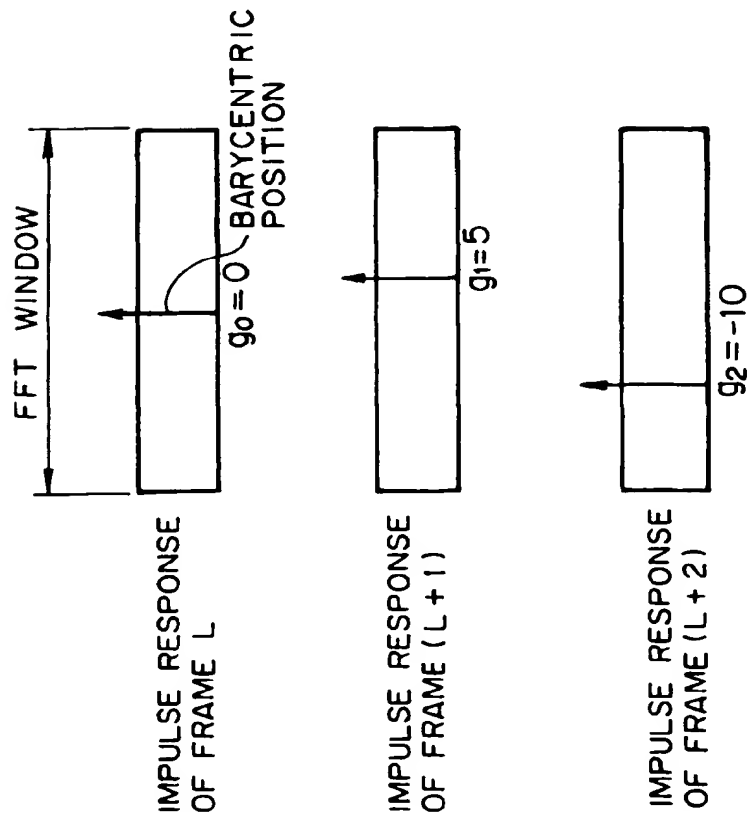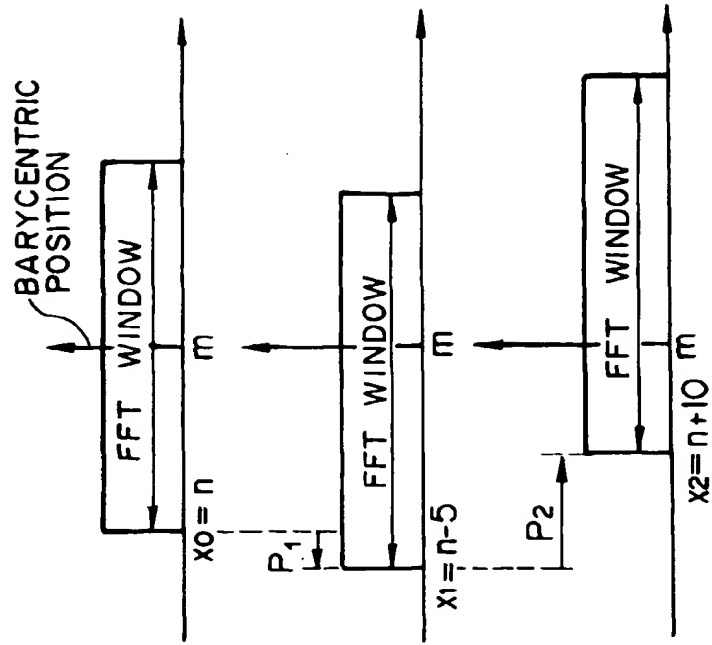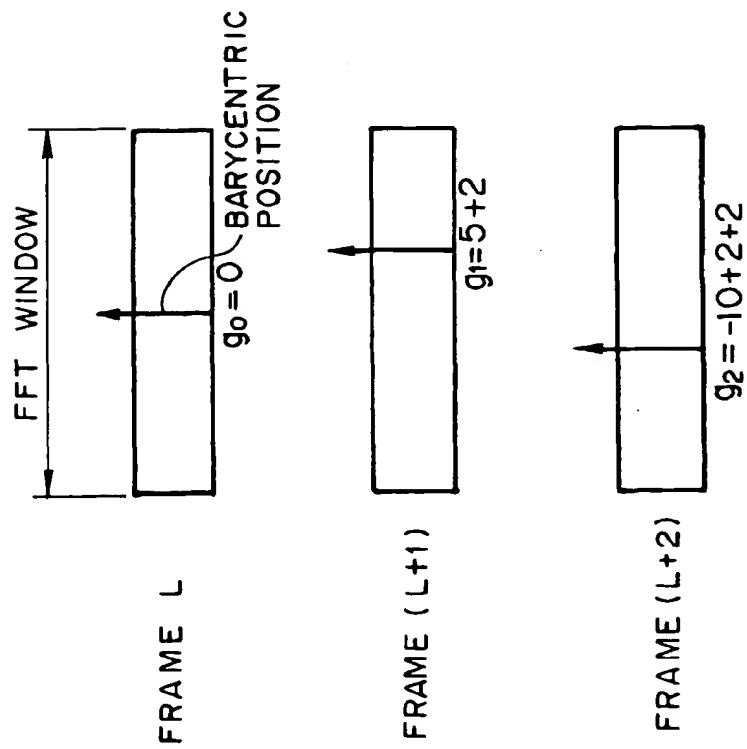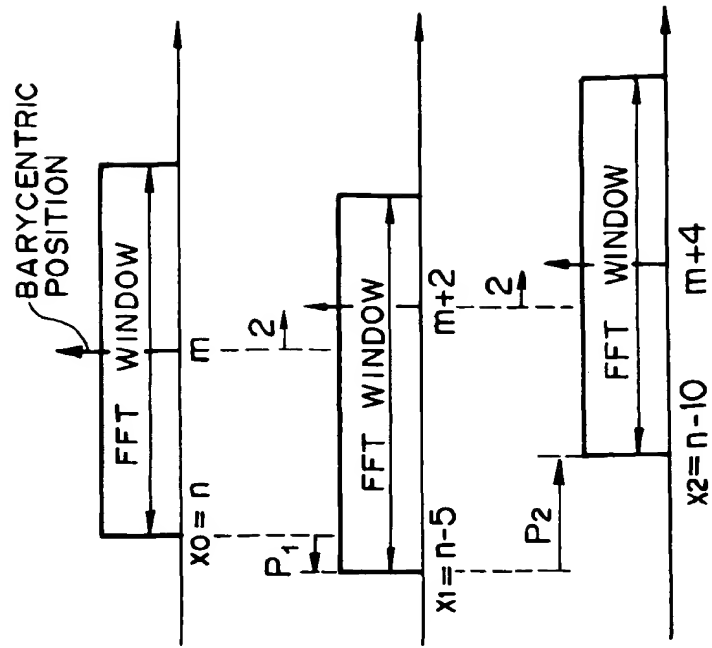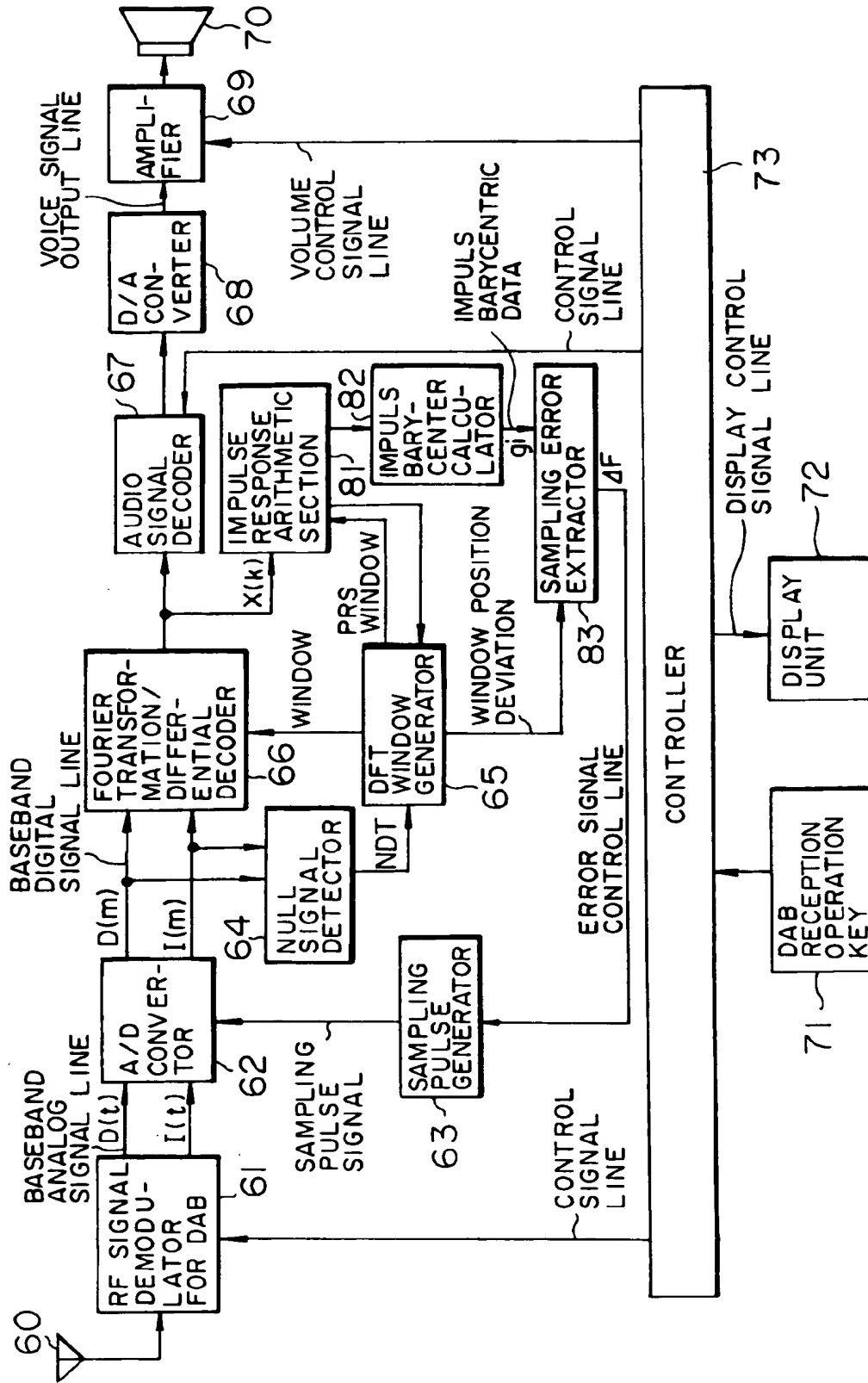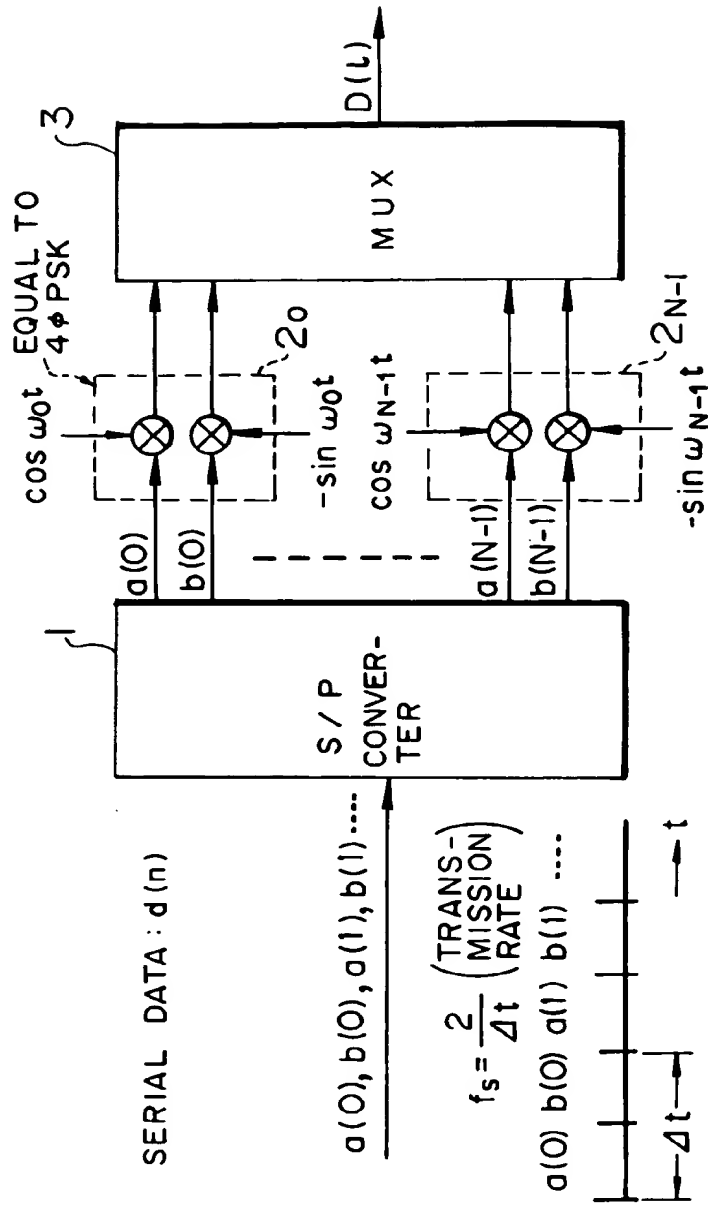
$$D(L) = \sum_{n=0}^{N-1} \{a(n) \cos \omega_n t + b(n) \sin \omega_n t\}$$

$$\begin{cases} \omega_n = 2\pi f_n \\ f_n = f_0 + n\Delta f \\ \Delta f = \dfrac{1}{N\Delta t} \end{cases}$$

FIG. 14
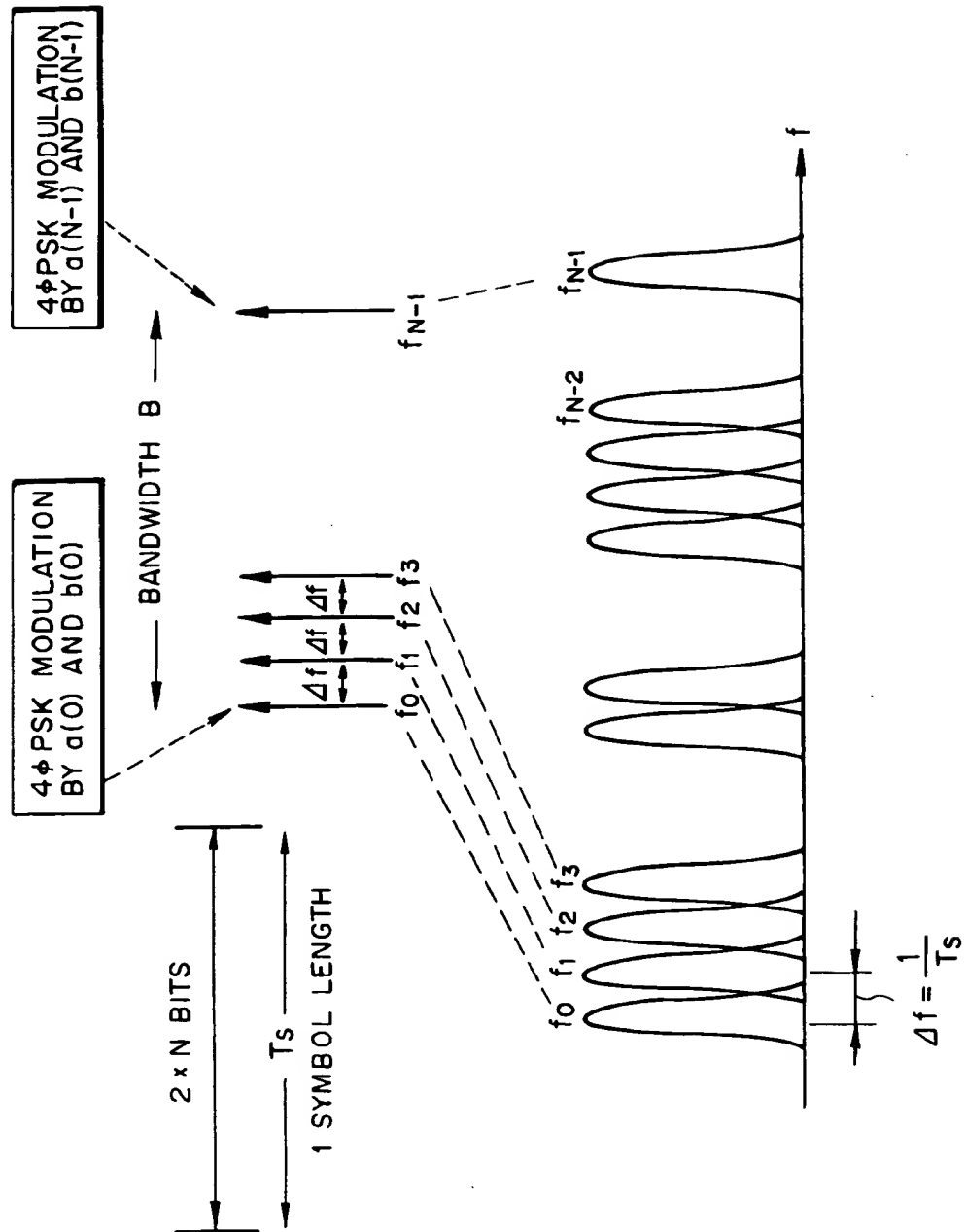
FIG.15

$$f_s = \frac{2}{\Delta t}$$
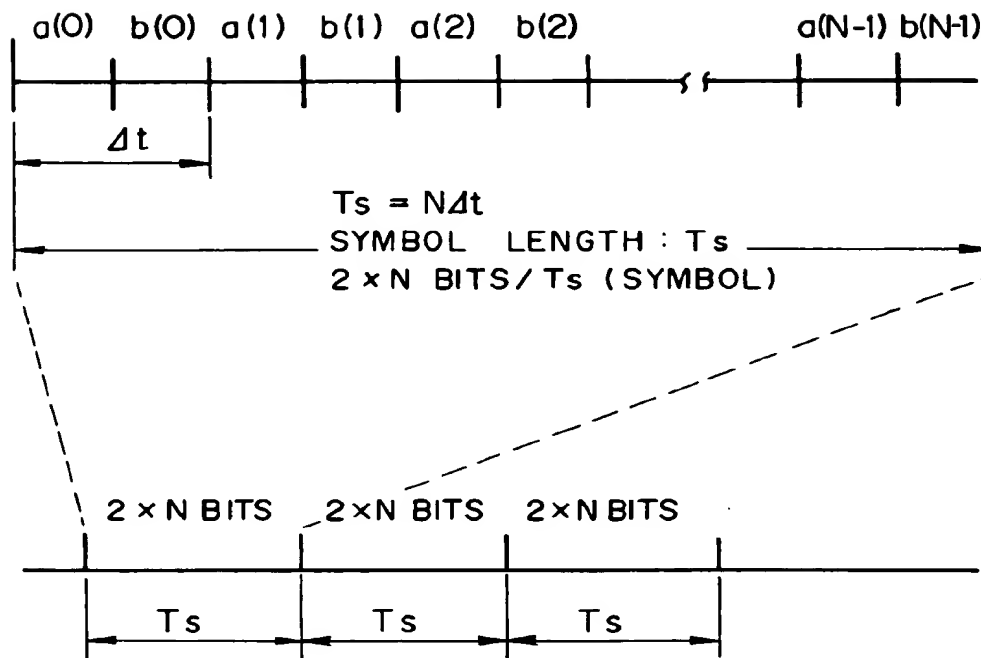
FIG. 16
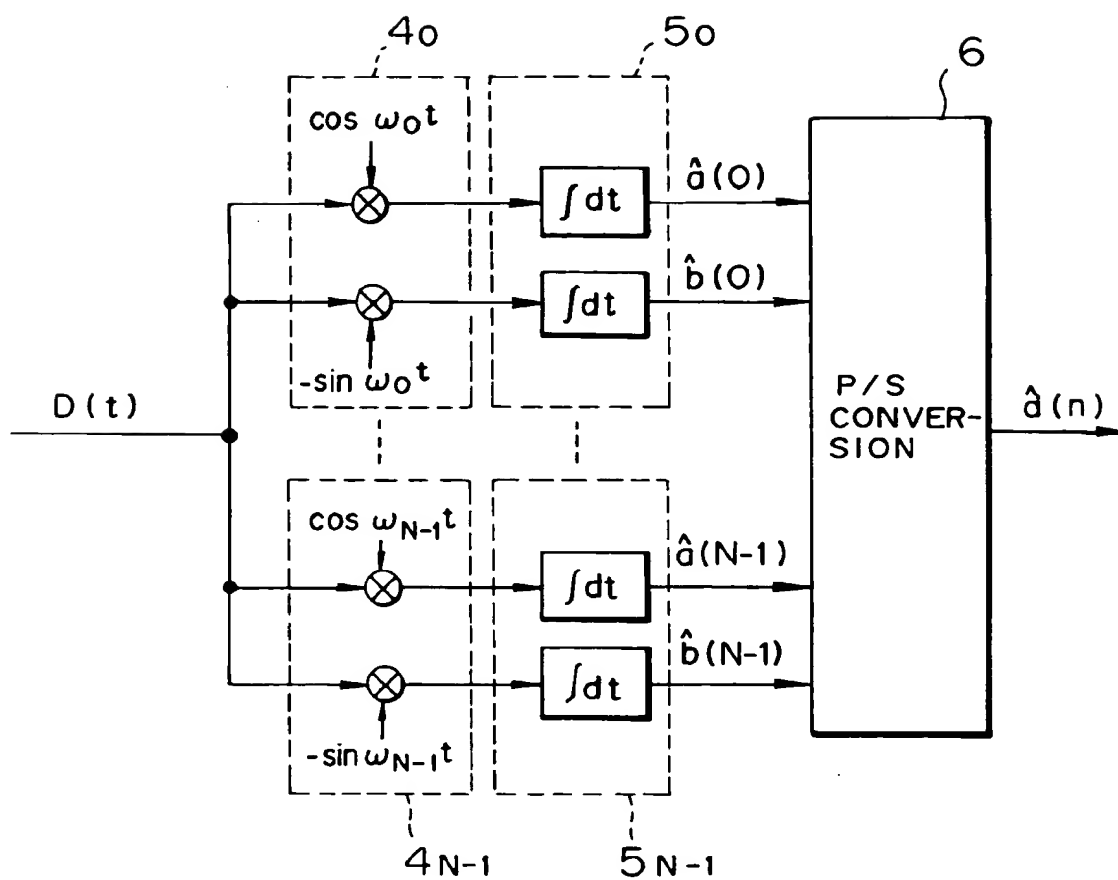
FIG. 17
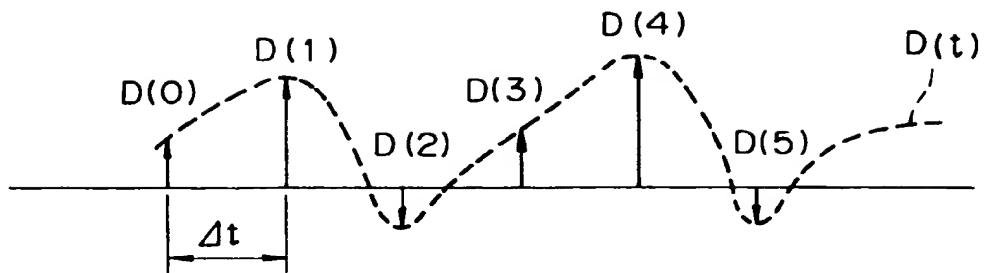
FIG. 18A

FIG. 18B

FIG. 19

FIG. 20A

D(t)

t

FIG. 20B

$N\Delta f$

f_0 f_{N-1}

FIG. 21

s(t) → ⊕ → r(t) → [FREQUENCY CONVERTER] 15

n(t)

$\hat{D}(t)$ → LPF 16a → A/D 17a → $\hat{D}(m)$

$\hat{I}(t)$ → LPF 16b → A/D 17b → $\hat{I}(m)$

DFT 18 → $\hat{d}(k)$

FIG. 22

REAL-
NUMBER
PART
D(m)

I(m)
IMAGINARY-
NUMBER PART

OFDM MODULATOR

IDFT

A/D 12a

A/D 12b

LPF 13a

LPF 13b

cos $\omega_c t$

D(t)

I(t)

14a

14b

H 14c

s(t)

sin $\omega_c t$

FREQUENCY
CONVERTER

14

11

FIG. 23

cos $2\pi f_c t$

s(t)

n(t)

r(t) 15a

BPF 15b

15c

15d

D'(t)

I'(t)

-sin $2\pi f_c t$

FIG. 24

$$d_l(k) = \frac{D_l(k) \cdot d_{l-1}(k)}{1+j}$$

DIFFERENTIAL ENCODER

21

$D_l(k) \longrightarrow$ [ ] $\longrightarrow d_l(k)$

$$D_l(k) = A_l(k) + jB_l(k) \qquad d_l(k) = a_l(k) + jb_l(k)$$

$D_0(k)$	$D_1(k)$	$D_2(k)$	$D_3(k)$	TRANSMISSION DATA
$d_0(k)$	$d_1(k)$	$d_2(k)$	$d_3(k)$	MODULATOR INPUT SIGNAL
$k=0 \sim N-1$	$k=0 \sim N-1$	$k=0 \sim N-1$	$k=0 \sim N-1$	
$l=0$	$l=1$	$l=2$	$l=3$	

$\longleftarrow Ts \longrightarrow$

FIG. 25

22

DEMODULATOR $\xrightarrow{d_l(k)}$

$$D_l(k) = (1+j) \frac{d_l(k)}{d_{l-1}(k)}$$

DIFFERENTIAL ENCODER

$\longrightarrow D_l(k)$

FIG. 26A

FIG. 26B

FIG. 27

DAB FRAME

NULL DETECTION SIGNAL

PRS WINDOW

DFT WINDOW

SAMPLING CLOCK

NULL GIT PRS

NULL PRS